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A Dimmable Resonant Inverter Electronic Ballast with Unity Power Factor

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Abstract — The conventional line-frequency magnetic ballast is heavy and bulky, which is too kept inside the compact fluorescent lamp (CFL). The paper describes about single stage dimmable electronic ballast with very high power factor and its high efficiency. A compact lamp power circuit is designed by using integrating a buck boost power factor corrector with a current-fed resonant inverter. Then the integration process gives a single power-processing unit that minimizes the number of circuit components. In this paper the proposed resonant inverter will reduces the circulating current in the resonant tank. It also allows simple gate drivers to be used so that isolation devices can be eliminated. The design, analysis and simulation were done using MATLAB SIMULINK.

Index Terms — *Electronic ballast, fluorescent lamps, power factor correction.*

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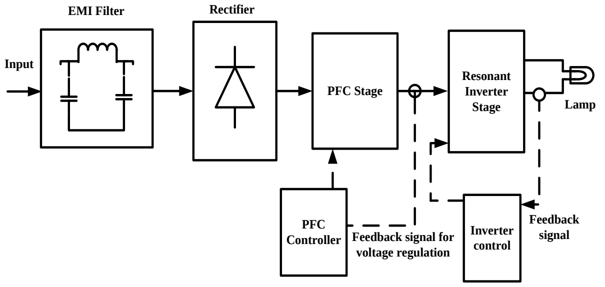
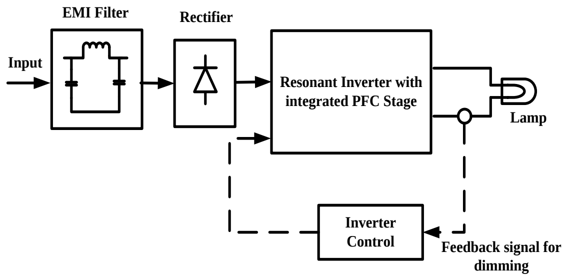
I. INTRODUCTION

In fluorescent lamps, ballast is required to provide sufficient high voltage for proper lamp ignition and stabilize the lamp current once the lamp arc is established. The conventional line-frequency magnetic ballast is heavy and bulky, which makes the device itself too large to be installed inside the CFL. The low operating frequency also causes light flickering, and it is impossible to implement dimming operations with magnetic ballast. A typical electronic ballast configuration consisting of a power factor correction (PFC) stage and a resonant inverter is shown in Fig. 1(a)[1]. This two-stage circuit [2] results in a high-cost and large-size circuit. Single-stage inverters (SSIs) are then proposed [3] by combining the PFC switch and one of the switches in the half-bridge resonant inverter together, as shown in Fig. 1(b), using the synchronous switch concept.

In this paper, a single-stage dimming electronic ballast with variable frequency control using an integrated buck-boost PFC current-fed resonant inverter is proposed. Although the efficiency of the SSI tends to be lower than the conventional design approach, the proposed SSI circuit saves one MOSFET and a PFC controller compared to the two-stage high-power-factor ballast circuit. It also allows a simple-switch driver circuit to be used without any isolation devices.



(a)



(b)

Fig.1. Typical electronic ballast systems. (a) Two-stages ballast circuit. (b) Single-stage electronic ballast.

II. CHARACTERISTICS OF THE PROPOSED CIRCUIT

A. Derivation of the Proposed Single-Stage Ballast Circuit

It is known that a boost converter is commonly used in PFC applications, as the boost inductor connected in series with the rectifier allows a continuous current to be drawn from the line [4]. However, when the boost converter switch is OFF and the boost diode conducts, the switch is connected directly to the output. Hence, the switch has to suffer a voltage stress of magnitude at least twice the rms value of the line voltage, assuming that a 50% duty ratio is used. The buck-boost converter is an alternative choice in PFC application. As the output voltage is not necessarily higher than the input voltage, it does not require a large-size high-voltage capacitor and a switch. Assume that the input voltage is a pure sinusoidal signal, as shown in (1), where V_p is the peak line voltage and f_L the line frequency. When the buck-boost PFC stage operates in discontinuous conduction mode (DCM) with a constant duty ratio, the peak of the DCM inductor current follows the sinusoidal envelope generated by the rectified line voltage. The average current ($I_{s,avg}$) in every switching period drawn from the line is shown in (2), where T_s is the switching period, D is the duty ratio, and L_b is the buck-boost inductor [5]. From (2), the average line current is proportional to the line voltage, and

ISSN: 2321-8169
3036 - 3041

hence, a unity power factor can be achieved. The input power is obtained by averaging the input power over one line-frequency cycle, as given in (3). The buck-boost inductor (*Lb*) can then be determined from (3). The voltage conversion ratio between the rectified voltage (*V*rect) and the voltage across *Cb* during DCM is given in (4), where *R* represents the load of the buck-boost converter [6].

$$V_s(t) = V_P \sin(2\pi f_L t) \cdots\cdots (1)$$

$$I_{s,avg}(t) = \frac{V_P D^2 T_S}{2L_b} \sin(2\pi fLt) \cdots\cdots (2)$$

$$Pin = \frac{1}{2\pi}\int_0^{2\pi} V_P \sin(2\pi fLt) I_{s,avg}(t)\, d(\omega_L t) = \frac{V_P d^2 T_S}{4L_b} \cdots\cdots (3)$$

$$\frac{V_{dc}}{V_{rect}} = -\frac{D}{\Delta_1} = -\frac{D}{\sqrt{2L_b / RT_s}} \cdots\cdots (4)$$

The derivation of the proposed single-stage ballast circuit is shown in Fig. 2. By using the synchronous switch concept outlined in the buck-boost switch *M* can be combined with the resonant inverter switch *M*1, as shown in the figure. As the source terminals of both MOSFETs are connected to ground directly, the isolation device for the high-side switch used in the conventional voltage-fed half-bridge resonant inverter is eliminated in the proposed design.

B . Principle of operation

Fig.3 shows the operating stages of the proposed circuit. The duty ratio is assumed to be operating at 50% for theoretical explanations. The working of the proposed circuit is shown for one switching cycle. There are three stages of operation. They are as follows:

Stage 1: *M*1 is ON and *M*2 is OFF, *D*in conducts and *I*in rises linearly. *D*1 also conducts and *IL*1 rises linearly at the inverter stage. The total switch current (*I*s1) during this stage is the sum of *IL*1 and *I*in.

Stage 2: *M*2 is ON and *M*1 is OFF, *D*in is OFF and *I*in decreases linearly through diode *Db*. Current *I*sn flows through *M*2 and resonates with the resonant tank.

Stage 3: *M*2 is still ON and *M*1 remains OFF, *I*in decreases to zero while *IL*1 continues to flow through the resonant circuit.

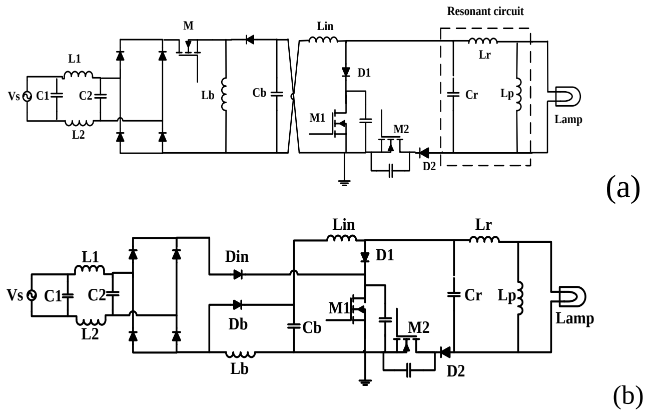


Fig. 2. Derivation of proposed ballast circuit. (a) Existing circuit (b) Derived circuit.

C. Characteristics of the Resonant Inverter

The resonant inverter stage employs a single input inductor (*L*in) current-source resonant inverter that consists of a resonant inductor (*Lr*) and a parallel capacitor (*Cr*) to form the corner frequency. The parallel inductor (*Lp*) is the starting inductor that provides sufficient high voltage to ignite the lamp. It also helps to reduce the circulating current in the resonant tank by providing more current to flow to the lamp at the desired switching frequency. Equation (6) gives the equation for the *Q*-factor of the proposed resonant circuit. By selecting high-enough *Q* value in the resonant circuit, close-to-sinusoidal waveforms are achieved at the output, and fundamental approximation can be applied .The characteristics impedance (*Zo*) of the resonant circuit is also defined, as in to relate the quality factor and lamp resistance.

$$f_0 = \frac{1}{2\pi\sqrt{L_r C_r}} \cdots\cdots (5)$$

$$Q = \frac{\omega_0 L_r}{R_{lamp}} \cdots\cdots (6)$$

$$Z_0 = \sqrt{\frac{L_r}{C_r}} = QR_{lamp} \cdots\cdots (7)$$

Before the lamp is ignited, the lamp resistance is infinite and the output of the resonant circuit can be modelled as an open circuit. The corresponding equivalent circuit during this phase is shown in Fig. 4. According to fundamental approximation the current *i*sn,1 (*t*) is the fundamental component of the square waveform input current and is given by (8), where *f*ph is the switching frequency during the preheat phase

$$i_{sn,1}(t) = I_{sn} \sin t\left(2\pi f_{ph} t\right) \cdots\cdots (8)$$

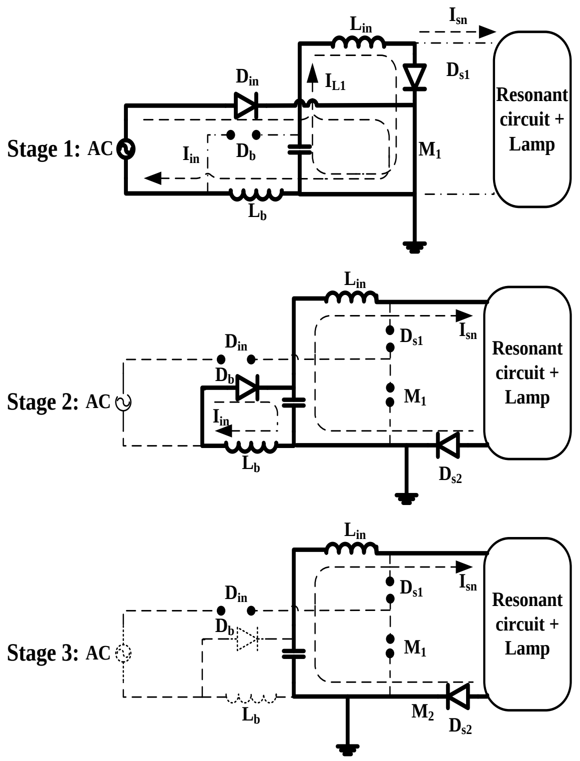


Fig.3. (a) Operating principles of the proposed circuit.

The current flowing through inductor *Lp* is obtained as given by using phasor calculation, with *L*eq = *Lr* + *Lp*. This current is also the preheat current flowing through the filaments during the preheat phase before the lamp is ignited. By solving (9) as a function of *f*ph, it is able to calculate the ballast preheat frequency (*f*ph) as shown in (10), which provides the proper preheat current through the lamp filaments. The output voltage across *Lp*, which is also the voltage across the lamp during the preheat phase. When $\omega = \omega$re, Where $\omega_{re} = 1/\sqrt{L_{eq} C_r}$, a very high voltage is achieved at the output that must be required for lamp ignition.

$$i_{ph}(\omega) = i_{sn,1} \left| \frac{1}{1 - \omega^2 L_{eq} C_r} \right| \cdots\cdots\cdots (9)$$

$$fph = \frac{1}{2\pi} \sqrt{\frac{1 - \left| i_{ph} / i_{sn,1} \right|}{L_{eq} C_r}} \cdots\cdots\cdots\cdots (10)$$

$$v_{Lp}(w) = i_{sn,1} \left| \frac{\omega L_p}{1 - \omega^2 L_{eq} C_r} \right| \cdots\cdots\cdots (11)$$

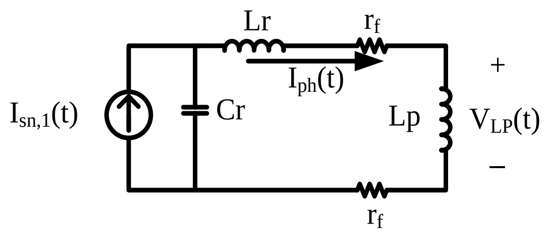


Fig. 4. Equivalent circuit before lamp ignition

After the lamp is ignited, the lamp resistance is no longer infinite. High output voltage is achieved when *Q* is low (i.e., lamp ignition) and low output voltage is obtained when *Q* increases as *R*lamp decreases after the lamp is ignited. Hence, the proposed resonant circuit is suitable in electronic ballast applications. Another feature of the proposed resonant circuit is that it naturally provides short-circuit protection when the lamp is accidentally shorted. The inductor *Lr* provides sufficient impedance to limit the current flowing through the shorted circuit path. The switching frequency of the resonant circuit should be decreased to lower the lamp power during dimming operations. Another feature of this resonant circuit is that the parameter *k* serves as another design parameter to adjust the resonant frequency from the corner frequency without changing the actual component values. According to eqn, *Lp* should be chosen to be larger than *Lr* s that sufficient high output voltage can be provided to start up the lamp.

III. Design Procedure

A 40W Lamp with a rated current of 0.2 Arms is chosen as the testing prototype. The design specifications are given as follows:

- Corner frequency(fo) 60KHz;
- Line frequency 230 Vrms, 50Hz;
- Quality factor (Q) 2.0-3.0

1) The full power steady state lamp resistance is calculated as given in (12), where *I*la,rms is the lamp rms current. The resonant circuit components *Lr* and
Cr are then calculated from (5) and (6), respectively,
by selecting *Q* to be 2.2.

$$R_{lamp} = \frac{P_{lamp}}{I_{la\ ,rms}^{2}} = \frac{40W}{(0.2A)^2} = 1000\Omega \ldots\ldots\ldots\ldots\ldots (12)$$

$$L_r = \frac{RlampQ}{\omega 0} = \frac{1000\Omega(2.2)}{2\pi * 60kHz} = 5.84mH \ldots\ldots (13)$$

$$Cr = \frac{1}{(2\pi * 60kHz)^2 (5.84mH)} = 1.2nF \ldots\ldots. (14)$$

ISSN: 2321-8169
3036 - 3041

2) *Lp* is chosen according to the *k* -value. In this design, *k* = 2; hence

$$L_p = Lrk = 5.84mh * 2 = 11.7mH \ldots\ldots\ldots\ldots(15)$$

3) The value of *Lb* can be calculated from (2) by substituting *D* = 0.5, *Ts* = 1/60 kHz, and *P*in = 40 W, assuming that the PFC stage efficiency (*η*) is 90%

$$L_b = \frac{V_p^{\,2} D^2 T_s}{4 P_{in} \eta} = \frac{(325)^2 0.5^2 (1/60kHz)}{4(40w)(0.9)}$$
$$= 3.06mH \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(16)$$

IV. SIMULATION RESULTS

The design, analysis and simulation is done using MATLAB SIMULINK version 8. Fig. 5 shows the input voltage and current to the ballast circuit. Fig.6 shows the output lamp voltage and lamp current waveforms. Fig.7 shows the envelope of the voltage and current across the lamp.

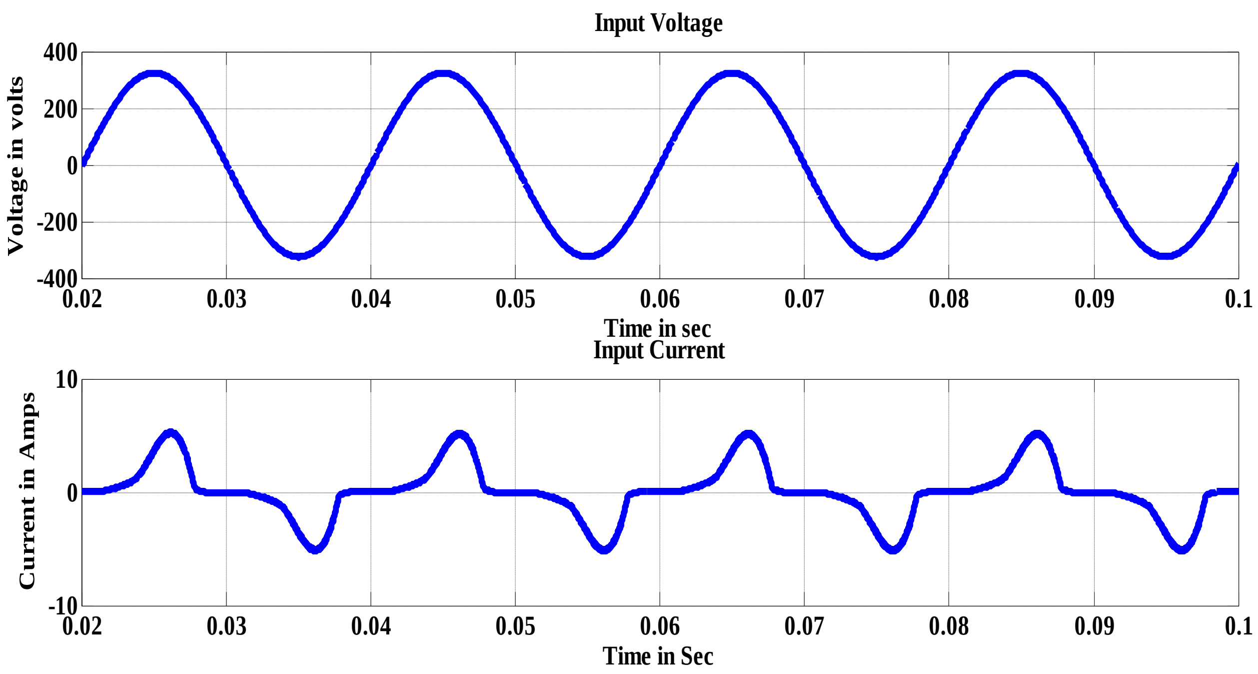


Fig.5. Input voltage and input current

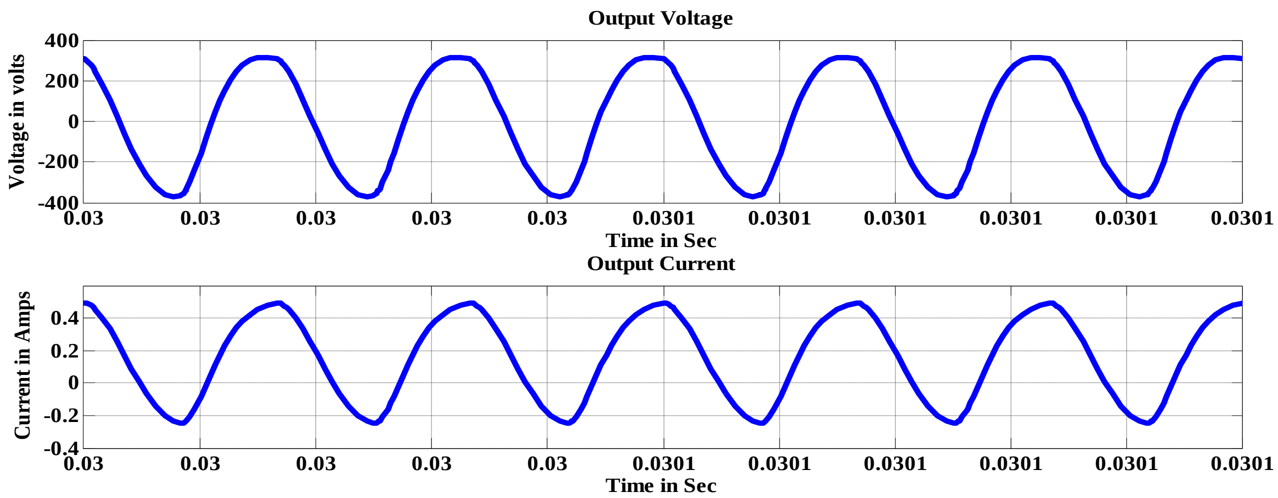


Fig.6. Lamp voltage and lamp current

ISSN: 2321-8169
3036 - 3041

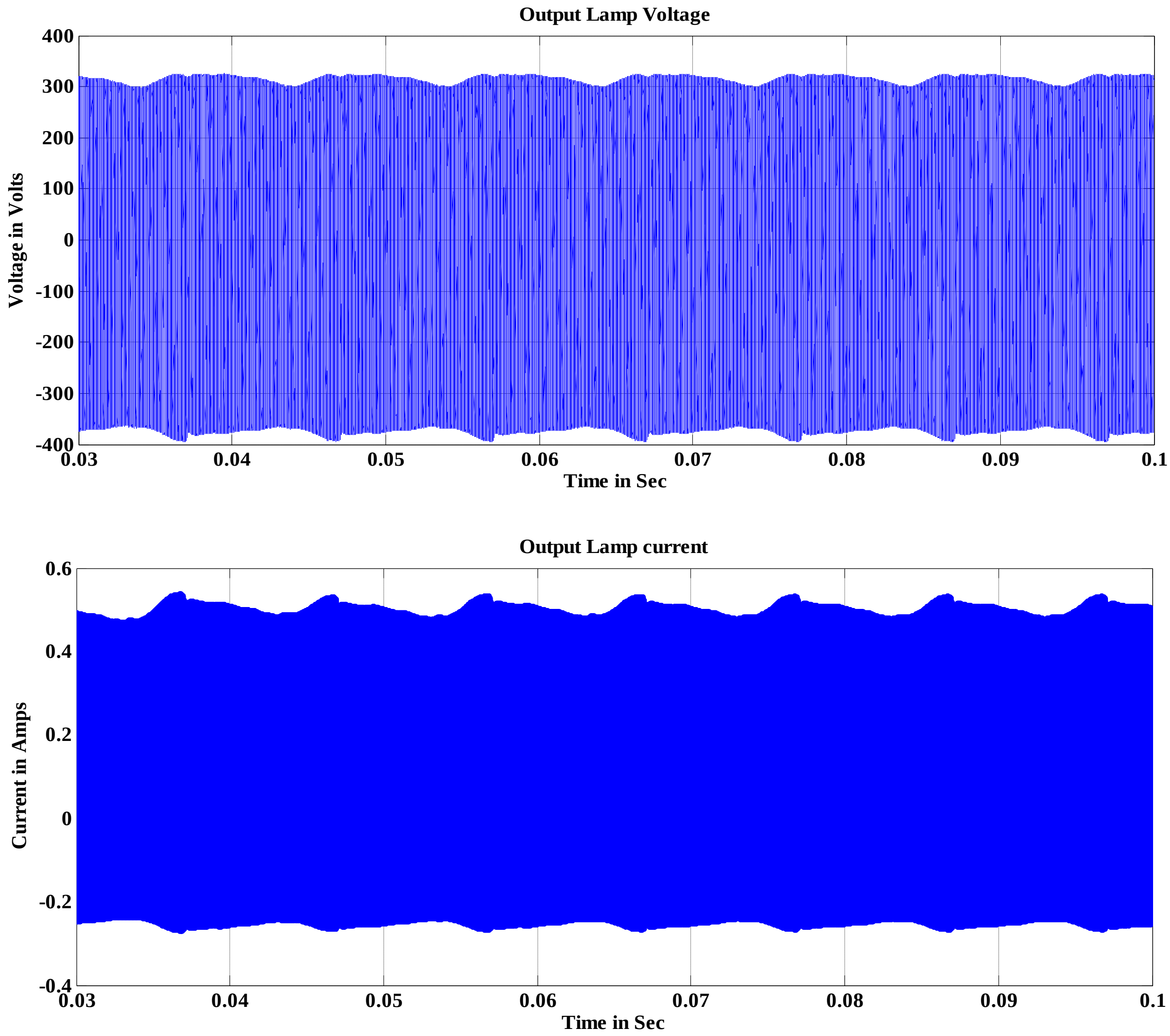


Fig. 7. Envelope of Output lamp voltage and lamp current

V. CONCLUSION

A new single-stage high-power-factor electronic ballast has been presented in this paper. It can be operated from 220V utility line. The proposed topology appears as a good solution to implement low cost high power factor electronic ballasts. A very high power factor (>0.995) and high efficiency (90.8%) has been presented.

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